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INFLUENCE OF HORIZONTAL GROUND WIRES ON LOW ANGLE RADIATION FROM HF ANTENNAS

C. L. Chen
W. L. Weeks

School of Electrical Engineering
Purdue University
West Lafayette, Indiana

Rome Air Development Center
Air Force Systems Command
Griffiss Air Force Base, New York 13441

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APPROVED:

CHARLES J. DRANE
Project Engineer

APPROVED:

ALLAN C. SCHELL, Chief
Electromagnetic Sciences Division

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20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

Digital computer programs are being developed to evaluate the effects of horizontal ground wire systems on the low angle radiation from vertically polarized HF antennas. An extensive method of moments program (WF-LLL2A) has been modified for this purpose. An alternative approximate method has also been studied and the results compared. Results to date indicate the

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feasibility of these types of calculations and that a low angle field enhancement of 20 to 25% by design of the ground wire system may be possible at the higher frequencies in the HF band.

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Summary

To make best use of ionospheric reflections for very long range transmissions of radio waves in the HF part of the spectrum, it is necessary to radiate vertically polarized waves having substantial field strength in the directions near the horizon. However, as is well known, the imperfect conductivity of typical earth tends to reduce drastically the available signal near the horizon. The possibility of increasing the radiation at such low angles by the interaction of a horizontal ground wire system is known and has been discussed but heretofore there has been very little quantitative information available on this subject and no systematic procedures for the design of such ground systems; in view of the substantial cost of such ground systems, much better information on their performance is needed.

The work described in this interim report has as its basic purpose the systematic determination of the effects that horizontal ground wire systems may have on the low angle radiation, and, eventually, the development of rational procedures for the design of such ground systems for application in specific tactical systems. In the course of this study, it has been demonstrated that useful results can be obtained with the aid of an extensive digital computer program incorporating the best available theoretical representation (Sommerfeld integrals) in a "moment method" for the determination of antenna currents and fields. It has also been shown (by comparison) that useful results may be obtained from a relatively simple digital computer program that, unfortunately, must be based on assumed current distributions in the ground wires.

The amount of low angle radiation enhancement that is feasible by design of such ground wire systems depends on earth conductivity and frequency. It ranges from just a few per cent at the lower end of the HF band to 25% or more at the high end of the HF band in a region having poor earth conductivity.

1. Introduction

For antennas operating in the HF range or lower, antenna performance suffers unless the typical poor earth conductivity found at most sites is enhanced by a system of conducting wires near or below the surface of the earth. This ground wire system is usually essential for impedance (i.e., lowered input resistance losses) control, but it might also be helpful in radiation pattern control, especially to alleviate the well known problem of the null at the horizon found with vertically polarized antennas. This report deals with only the latter problem; it describes investigations into the extent to which a ground wire system may be used to enhance the radiation pattern of some simple vertically polarized arrays.

Evaluation of the effects of a ground screen on or in an actual earth by standard boundary value solutions of the field equations is impractical at the present time. However, over the past 10 to 12 years, a method of calculation of current distribution, impedance, radiation and scattering patterns of wire like structures has been highly developed. A number of computer programs, based on the method (that has come to be known as the method of moments) have been developed to handle wire-like structures of considerable complexity and variety. In the main, the programs treat the antennas acting only in free space, or at most, in the vicinity of a perfectly conducting ground. A group at Lawrence Livermore Laboratories, however, modified and extended a program[1] of this type (WAMP) to cover wire structures in or near a plane homogeneous earth having arbitrary conductivity. They called the program WF-LLL2A and described it in the "Electromagnetic Computer Code Newsletter" Volume 2 No. 1 dated April 1, 1975 and in UCRL-51821[2]

"Fortran Subrountines for the Numerical Integration of Sommerfeld Integrals Unter Anderem", by Lager and Lytle, dated May 21, 1975. A group at Purdue had been pursuing a similar development and upon learning of the Livermore work decided to try to adopt and adapt the WF-LLL2A program for the purpose of doing ground screen calculations, although neither the WF-LLL2A program nor the modifications of it had yet been tested or evaluated for the purpose.

But the basic idea was to restrict potential ground screen systems to wire structures, and to regard the wires in the ground screen as part of the antenna in or near the earth. By and large, this meant that the fields due to the elemental currents flowing in the antenna wires had to be evaluated with the Sommerfeld integrals, or suitable approximations to them. It turned out that none of the existing codes would converge properly when the wire elements were close to the earth and to each other. Locating the source of the difficulty and correcting it involved considerably more time and effort than had been anticipated.

2. Background of the Wire Antenna Program

The computer programs assume the structure consists of thin wires and essentially solve an electric field integral equation, numerically, for the currents that result with a stated applied electric field (or voltage). The program WF-LLL2A employs a three term current expansion function for the current in each subsection of wires.

The essence of the method is to replace an integral equation for the electric field by a set of linear equations relating the electric field at a set of discrete points to a set of current levels that produce the fields at the set of points. The coefficients in the set of linear equations have the character of impedance coefficients and the matrix relating the electric fields and current levels is often called the impedance matrix, or system matrix or system impedance matrix. The task here is two fold 1) first compute the elements of the system impedance matrix, which means calculate the field due to a unit amplitude current segment in the appropriate environment at each of the field points 2) having obtained the impedance matrix, solve the resulting set of linear equations for the currents (i.e., invert the matrix).

The difficulty of applying this method to the calculation of the effects of wire ground screens on antenna arrays is two-fold. First, it is difficult to calculate the field produced by current elements at nearby points when the current element is near the surface of an imperfectly conducting earth. Second, with the ground wire system plus the antenna wires the number of segments (i.e., the number of equations in the linear set) into which the wire configurations must be broken is very large. Thus, the basic problem is one of time and storage space in the digital computer.

The WF-LLL2A has available within it four methods for the calculation of the fields in the presence of ground. Two of these are based on plane wave reflection coefficients to account for the reflected fields. The fields that would result from a perfectly conducting ground (i.e., an image) are modified according to the actual conductivity and dielectric constant of the ground. A third method makes use of the well known formulas developed by Norton; this method is used when the distances and ground parameters are in the proper range for the Norton formulas. The fourth method incorporated is the numerical evaluation of the Sommerfeld integrals, and for this there are two options available, the Hankel function form and the Bessel function form, with the choice between the two being made on the basis of position and parameter values. The integrals are evaluated by performing a contour integration in the complex plane, numerically, using an adaptive Romberg integration.

3. Modifications for Calculation of Effects of Ground Wire Systems

In the original program, for points nearby thin wires located close to the surface of the earth, the Sommerfeld integral calculations referred to above are extremely time consuming, if indeed they converge properly at all.

It was recognized from the outset that computer time would be a problem, so the program was reorganized in several respects. First, since the most time consuming calculations are nearby "self" and "mutual" impedance coefficients, the matrix was partitioned so that these time consuming calculations, which it was observed, actually need to be done only once for a given size wire, frequency, and ground parameters set, could be calculated once and for all and fed in thereafter, when successive runs were being made for the purpose of evaluation of the effects of different sizes and shapes of ground screens. Other programming techniques were developed so as to conserve the space available in memory by having in the fast memory only those parts of the program needed for the particular calculation being done.

A basic difficulty was encountered, however, in the use of the Sommerfeld integral portion of the program, when it was discovered that for interesting wire sizes, ground parameters and positions, a time of 1048 seconds in the CDC 6500 computer (one of the "priority" time limits set by The Purdue Computing Center) was insufficient to calculate the self impedance coefficients for even a very simple single wire structure. Moreover, the indications were that even a significantly larger amount of time would also be insufficient. It appeared then that some of our modificiations to the calculation would be essential and a step-by-step replacement and checking of various parts of the existing Sommerfeld integration packages was undertaken. Suffice it to say, that although most of the modifications decreased the time necessary for the calculation, they did not decrease it enough to be really useful and even so, some were giving up accuracy in the calculation. A very considerable amount of time was spent in tracking down the real source of the difficulty, and in trying to determine the accuracy of the modified systems.

To explain the difficulty involved in carrying out the numerical integration, consider the expression for the radial component of the electric field in the medium above the ground due to a horizontal electric dipole moment p oriented along the x-axis with notation as in the WF-LLL2A program:

$$E_\rho^H = \underbrace{\cos\phi}_{\text{GFIELDS}} \underbrace{\left[\frac{-j\omega p \mu_o}{4\pi k_2^2} \left\{ \left(\frac{\partial^2}{\partial \rho^2} + k_2^2 \right) (G_{22} - G_{21}) \right\} \right]}_{\text{GEVALA}}$$

$$+ \underbrace{\cos\phi}_{\text{SFIELDS}} \underbrace{\left[\frac{-j\omega p \mu_o}{4\pi k_2^2} \left\{ \frac{\partial^2}{\partial \rho^2} (k_2^2 V_{22}) + k_2^2 U_{22} \right\} \right]}_{\text{EVALUA2, EVALUA3}} \tag{1}$$

where

$$G_{22} = \frac{e^{-jk_2 (\rho^2 + (h-z)^2)^{\frac{1}{2}}}}{(\rho^2 + (h-z)^2)^{\frac{1}{2}}} \tag{2}$$

$$G_{21} = \frac{e^{-jk_2 (\rho^2 + (h+z)^2)^{\frac{1}{2}}}}{(\rho^2 + (h+z)^2)^{\frac{1}{2}}} \tag{3}$$

$$V_{22} = \int_{-\infty}^{\infty} \frac{e^{-\gamma_2 (h+z)}}{k_1^2 \gamma_2 + k_2^2 \gamma_1} H_0^{(2)}(\lambda\rho)\lambda \, d\lambda \tag{4}$$

$$U_{22} = \int_{-\infty}^{\infty} \frac{e^{-\gamma_2(h+z)}}{\gamma_1 + \gamma_2} H_0^{(2)}(\lambda p)\lambda \, d\lambda \qquad (5)$$

and $k_1 = [\omega^2\mu_o\varepsilon_o\varepsilon_{r1} - j\omega\mu_o\sigma_1]^{\frac{1}{2}}$, $k_2 = \omega\sqrt{\mu_o\varepsilon_0}$ are, respectively, the complex propagation constants for medium 1 (ground) and 2 (air); and $\gamma_1 = [\lambda^2 - k_1^2]^{\frac{1}{2}}$, $\gamma_2 = [\lambda^2 - k_2^2]^{\frac{1}{2}}$. [To avoid possible confusion, we have assumed $e^{j\omega t}$ type of time dependence. This change dictates the use of the Hankel function of the second kind and a different choice in branch cuts. All other notations are the same as those of Lager and Lytle.]

To evaluate the elements of the impedance matrix, the tangential component of the electric field at the surface of the wire is integrated with respect to the dipole moments situated along the center of the wire. For example, consider the self impedance of a segment of wire of radius a and length ℓ. The field at x=0, y=a, z=h, is integrated along a path described by the conditions $-\frac{\ell}{2} < x < \frac{\ell}{2}$, y=0, z=h, as indicated in Fig. 1.

First, focus attention to the first part of the expression (GFIELDS and GEVALA). When the wire is several radii above ground, h>>a, G_{21} is a relatively smooth function of x while G_{22} has a peak at x=0 with a peak value of the order of a^{-1}. When the wire is close to the ground, h~a, both G_{21} and G_{22} have peaks at x≈0, and $\frac{d^2}{d\rho^2}(G_{22} - G_{21})$ has a sharp peak with a peak value of the order of a^{-5}. In the original WF-LLL2A program, the integration is performed via an adaptive Romberg method. Due to the sharp peak, the integration is extremely time consuming for thin wires close to ground. Actually, the sharp peak may be smoothed by partial integration. We replaced this portion of program (GFIELDS and GEVALA) by two closed form expressions and an integration of an integrand with a moderate peak.

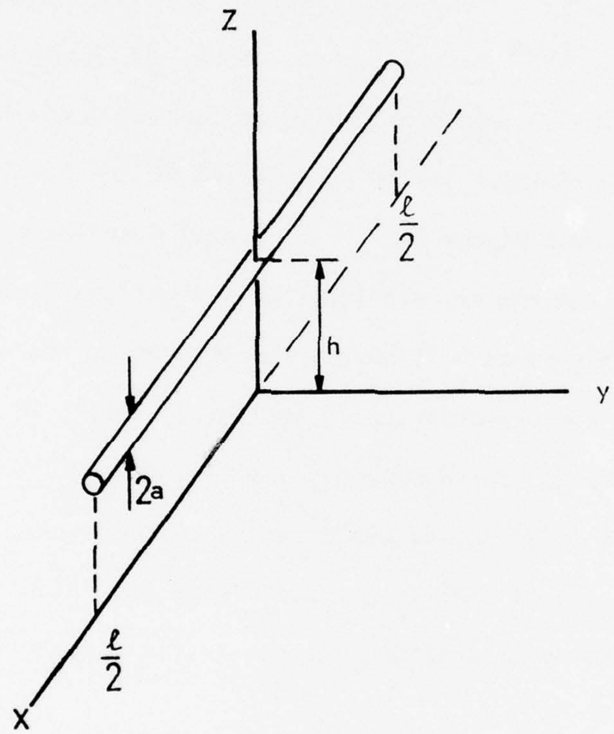


FIG. 1 GEOMETRY OF WIRE ELEMENT

Now consider the second part of equation (1) (SFIELDS and EVALUA2 (or EVALUA3)). Double integrals are involved in calculating the parts of this contribution to the impedance matrix, and the evaluation of V_{22} and U_{22} (equations(4) and (5)) is particularly time consuming. For $|h+z| \geq |\lambda|$, the exponential functions in (4) and (5) decay rapidly and thus V_{22} and U_{22} may be evaluated routinely. However, when $|h+z| << |\lambda|$, the integrands of (4) and (5) behave like $|x|^{\frac{1}{2}} e^{-j|\lambda|\rho}$ for $|\lambda\rho| >> 1$. Thus, the integrands decay extremely slowly and the integration has to be carried out at least to $|\lambda| >> \rho^{-1}$. In addition, the contour has to be deformed in the complex λ-plane so as to avoid the poles and branch cuts. The Hankel function $H_o^{(2)}(\lambda\rho)$ with complex argument $\lambda\rho$ is extremely difficult to evaluate particularly for the case $\rho << 1$. Since our concern is mainly for the region where $\rho << 1$, $h+z << 1$, which means that the time retardation is negligibly small, we replace (4) and (5) by their quasi-static approximations, i.e., $\gamma_1 \sim \lambda$, $\gamma_2 \sim \lambda$. With such approximations, (4) and (5) may be evaluated in closed forms. Our calculation shows that in the region of interest, these closed-form expressions are very accurate. In our version of SFIELDS, the quasi-static approximation is used for $\rho \leq 5a$.

In these ways we reduce the time required for self and nearby mutual impedance terms to manageable proportions and have thus arrived at our "best method to date". We will refer to this modified program as WF-LLL2AP.

4. Alternative Method of Calculation

An approximate but very fast method of evaluating the effect of ground wires on the radiation patterns of vertical antennas over actual grounds may be based on the reciprocity theorem. Fig. 2 shows an antenna array on an imperfect earth. The current distribution in the array is defined to be J and is assumed known (in most cases a simple sinusoidal distribution along the wires is a

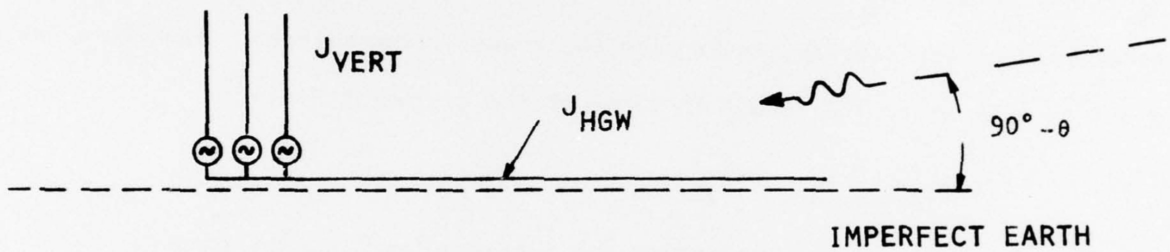


FIG. 2 RADIATION AT LOW ANGLES INCIDENT ON ANTENNA ARRAY OVER IMPERFECT GROUND

sufficiently good approximation). The open circuit voltage at this input of the array can be found from the integral

$$V_{oc}\ (\theta,\phi) = \int_{array} J \cdot E\ dv$$

assuming a unit current at the input on transmitting. In this expression E is the electric field of a plane wave arriving at angle (θ,ϕ) in the presence of the imperfect earth (and of course evaluated at the positions of the currents in the elements of the array), but in the absence of the array. $V_{oc}\ (\theta,\phi)$ is then the receiving voltage pattern of the array over an imperfect ground and by reciprocity, it is also the transmitting pattern. Therefore, we might as well write the result in terms of the distant E field[3*]

$$E(\theta,\phi) = \int E_{inc} \cdot J_t\ dv$$

where J_t is the current distribution (as a transmitter) on the antenna system when it is transmitting. The advantage of this formulation is that except for grazing angles, the field E_{inc} may be determined with plane wave reflection coefficients (for grazing angles the Sommerfeld solutions or equivalent should be used).

Moreover, if the ground system consists of a set of ground wires, it is both consistent and practical to regard the currents in the ground wires as a part of the transmitting antenna current distribution, that is

$$J_t = J_{vert} + J_{HGW}$$

where the two terms are, respectively, the current density in the vertical elements and (J_{HGW}) the current density in the horizontal ground wire. Then the radiated field is

$$E(\theta,\phi) = \int E_{inc} \cdot J_{vert}\ dv + \int E_{inc} \cdot J_{HGW}\ dv$$

---

*(Weeks, W. L., Antenna Engineering, pg. 24 equation 23)

It is to be expected that J_{vert} will not change significantly in distribution whether there is a ground wire system or not, so with unit current at the input in both cases:

$$E(\theta,\phi)]_{\text{no ground screen}} = \int E_{inc} \cdot J_{vert}\, dv$$

Thus, it is possible to define an enhancement of the radiated field due to the presence of the ground wire system as

$$\frac{E(\theta,\phi)}{E(\theta,\phi)]_{\text{no ground screen}}} = 1 + \frac{\int E_{inc} \cdot J_{HGW}\, dv}{\int E_{inc} \cdot J_{vert}\, dv}$$

For example, if the ground is perfect and the ground system wires lie on the perfect ground, then $E_{inc} \cdot J_{HGW} = 0$ and no enhancement is possible. In this case

$$E(\theta,\phi) = \int E_{inc} \cdot J_{vert}\, dv$$

is the pattern of the transmitting antenna array over a perfect ground.

If we specify that the elements are along the vertical direction z and the plane wave is incident at angle θ in the x-z plane with H perpendicular to the plane of incidence, the E_{inc} field has a form as follows (in the air above the ground)

$$E_{z,inc} = \frac{\beta_x}{\omega\, \varepsilon_0} H_0\, e^{j\beta_x x} [e^{j\beta_z z} + \rho^{H1}(\theta)\, e^{-j\beta_z z}]$$

$$E_x = \frac{-\beta_z}{\omega\, \varepsilon_o} H_0\, e^{j\beta_x x} [e^{j\beta_z z} - \rho^{H1}(\theta)\, e^{-j\beta_z z}]$$

where $\beta_x = \beta_0 \sin\theta$ and $\beta_z = \beta_0 \cos\theta$ and $\beta_0 = \omega\sqrt{\mu_0\varepsilon_0}$; $\rho^{H1}(\theta) = \frac{H_{ref}}{H_{inc}}$ with wave incident at angle θ. In the ground (earth)

$$E_x = \frac{-\gamma_{ze}}{j\omega\varepsilon_e} (1 + \rho^{H1}(\theta))e^{\gamma_{ze} z + \gamma_{xe} x}$$

where z is the depth to be inserted as a negative number, and

$$E_z = \frac{-\gamma_{xe}}{j\omega\varepsilon_e}\,[1 + \rho^{H1}(\theta)]_e\ \gamma_{ze}z + \gamma_{xe}x$$

$$\gamma_{xe} = j\beta_x = j\beta_0 \sin\theta$$

$$\gamma_{ze} = \sqrt{-\omega^2\mu_e\varepsilon_e - \gamma_{xe}^2} = \beta_0\sqrt{-\mu_{re}\varepsilon_{re} + \sin^2\theta}$$

$$\varepsilon_{re} = \varepsilon_G^1 - j\frac{\sigma}{\omega} \text{ so that } \gamma_{ze} \simeq \gamma_e = \sqrt{-\omega^2\mu_e\varepsilon_e}.$$

If the ground wire is at or very near the surface, z=0, only the component E_x is significant to the ground wire calculation and the coordinate value z is essentially zero.

All of the complex quantities in the expression above are easily determined with the aid of a digital computer. Only the quantity J_{HGW} is troublesome to obtain. Again, however, it is to be expected that on a ground wire of length L, the current on the wire would have the form

$$I_{HGW}(L) = A\,e^{\gamma_w x} + Be^{-\gamma_w x}$$

where γ_w is a propagation constant of the fields or current along the wire in the environment (i.e., near or in the earth and perhaps insulated). The value of γ_w is difficult to determine exactly analytically. Consequently, the enhancement was examined for different values of γ_w. For a thin wire exactly on the surface, one result (due to Coleman[4]) is

$$\gamma_w = j\frac{\beta_0}{\sqrt{2}}\sqrt{\frac{\varepsilon_G^1}{\varepsilon_0} + 1 - \frac{j\sigma}{\omega\,\varepsilon_0}} \tag{6}$$

where ε_G^1 is the real part of the earth permittivity, σ is the earth conductivity and β_0 is the free space propagation constant.

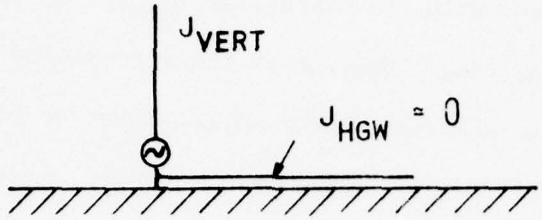
Generally speaking, for insulated wires, the attenuation will be less than the Coleman result and the phase constant will be more like that in free space. The net result is that the enhancement with such insulated wires is greater.

The quantities A and B in the expression $I_{HGW}(L)$ are determined primarily by the terminations. However if the attenuation is large, only one of the constants is important (B,for wires extending toward positive x).

In relating the current in the horizontal ground wires to the current at the input (or reference point) of the vertical antenna, a difficult, ticklish, but very practical consideration arises. For the field enhancement due to the ground wires certainly depends on the relative level of the currents in the ground wires; yet the level relative to the current in the verticals depends on the minute details of the method of attachment of the transmitting structure to the ground and to the ground wires at the point of feed. The significance of this is suggested by the sequence of sketches in Fig. 3. Sketch a suggests a limiting situation in which no current at all would flow in the ground wire. The transmitter in a is connected both to a perfect ground and to a slightly elevated ground wire, in this case, essentially all of the current would flow into the perfect ground. On the other hand, another extreme is indicated in sketch b in which there is no direct connection to the ground. Although it might seem at first sight that all of the current would flow into the slightly elevated ground wire, this would not in fact happen for there would be some hard to define capacitive susceptance to ground which would make the ground wire current somewhat less than the vertical input current, as suggested in sketch c. Sketch d shows something more like an actual situation with a ground connection to the imperfect earth as well as to the horizontal ground wires. In this case, the impedance to earth will be significant, complex, and sensitive to detail; and the input current at the vertical will thus divide in some (hard to determine) fashion between the horizontal ground wires and the grounding connection. From the point of view of trying to maximize the enhancement of fields by the ground wires,

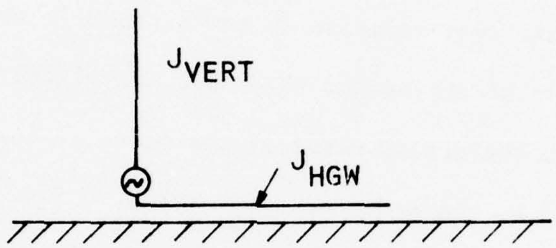
FIG. 3 MODELS ILLUSTRATING THE FACTORS THAT INFLUENCE THE RELATIVE LEVEL OF CURRENT IN THE GROUND WIRES

A.



PERFECT GROUND

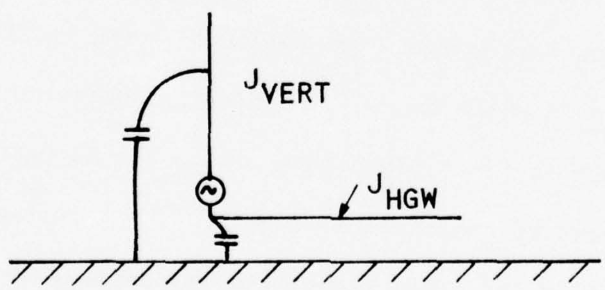
B.



NO DIRECT GROUND CONNECTION

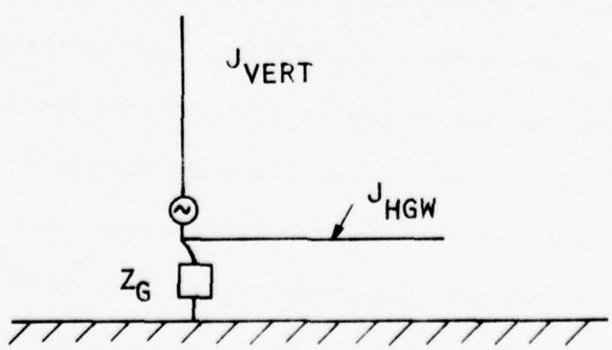
PERFECT GROUND

C.



MORE REALISTIC MODEL OF SITUATION DEPICTED IN PART B

D.



Z_G MODELS THE IMPEDANCE OF THE CONNECTION TO A POOR GROUND it is necessary that the ground wires have all of the current. This implies the highest possible grounding impedance. Yet other considerations (safety and lightning for example) require that the impedance to ground be small. This suggests that a technique for grounding which has a high impedance at HF but a low impedance at lower frequencies should be a worthwhile step to take in a system design in which the ground wires are being called upon to enhance the radiation pattern.

Another practical consideration that influences the enhancement possible by a ground wire system of a given size is the manner of termination of the wires(really, the terminating impedance of the ground wires at the ends remote from the transmitter). This termination impedance may be low (if connected to a low impedance ground electrode) or high (if the ends are capped with insulating material) or even "matched", if enough is known about the environment of the ground wires. Depending on the actual terminating impedance, of course the "input impedance" to the ground wire system viewed from the transmitter is different, and therefore a different relative level of current actually flows into the ground wires. In the case of earth having low conductivities, certain lengths would exhibit resonance type effects which would make the effective input impedance to the ground wires quite high (for example, half-wavelength wires left open ended, or quarter wavelength wires having low impedance terminations)

All of these effects make it difficult to determine what the levels of current in the ground wires are relative to those in the vertical wires. Thus, except for displaying the sensitivity of the low angle radiation enhancement to such effects, for the most part in the approximate calculation, we will assume a traveling wave distribution on the ground wire, with equal current levels flowing into the verticals and to the horizontal ground wire system.

The output of the computer program WF-LLL2AP can give guidance as to which assumptions or conditions are likely to be the better.

5. Results of the Calculations

The overall effects that radiating ground wires have on vertical radiation is suggested by the radiation patterns displayed in Figs. 4 and 5. These patterns were actually obtained by the approximate method described above, but Figs. 6 and 7 show similar results obtained by the modified WF-LLL2AP program and the results are then expected to be typical. The patterns show $E_\theta(\theta)$ in the forward direction of the vertical plane. In each case, the lower graph shows the pattern of the vertical array with no ground wires at all. The figures display the tendency of the field to have a null on the horizon (θ=90°) as expected; Fig. 4 gives the result for earth conductivity 10 millimhos/m, Fig. 5 gives those for 1 millimho/m. The upper curves in each case give the pattern with a ground wire system consisting of four horizontal ground wires at the surface of the earth in the forward direction only, extending to a length of 6 m (at 10 MHz). (The specific results depend on the ground wire length, and as will be seen below; longer ground wires do not necessarily provide more low angle field enhancement). Both of the figures show that ground wires enhance the radiation at all angles, assuming that the current in the ground wires near the input is equal to that in the vertical wires, and that the current distribution in the ground wires is that of a pure traveling wave at a propagation constant as calculated by Coleman's formula. It is also seen that the enhancement is greater with the lower earth conductivity, as might be expected.

The radiation patterns are all changing so rapidly at low angles (θ near 90°) that the positive effects of the ground wires, if any, are hard to

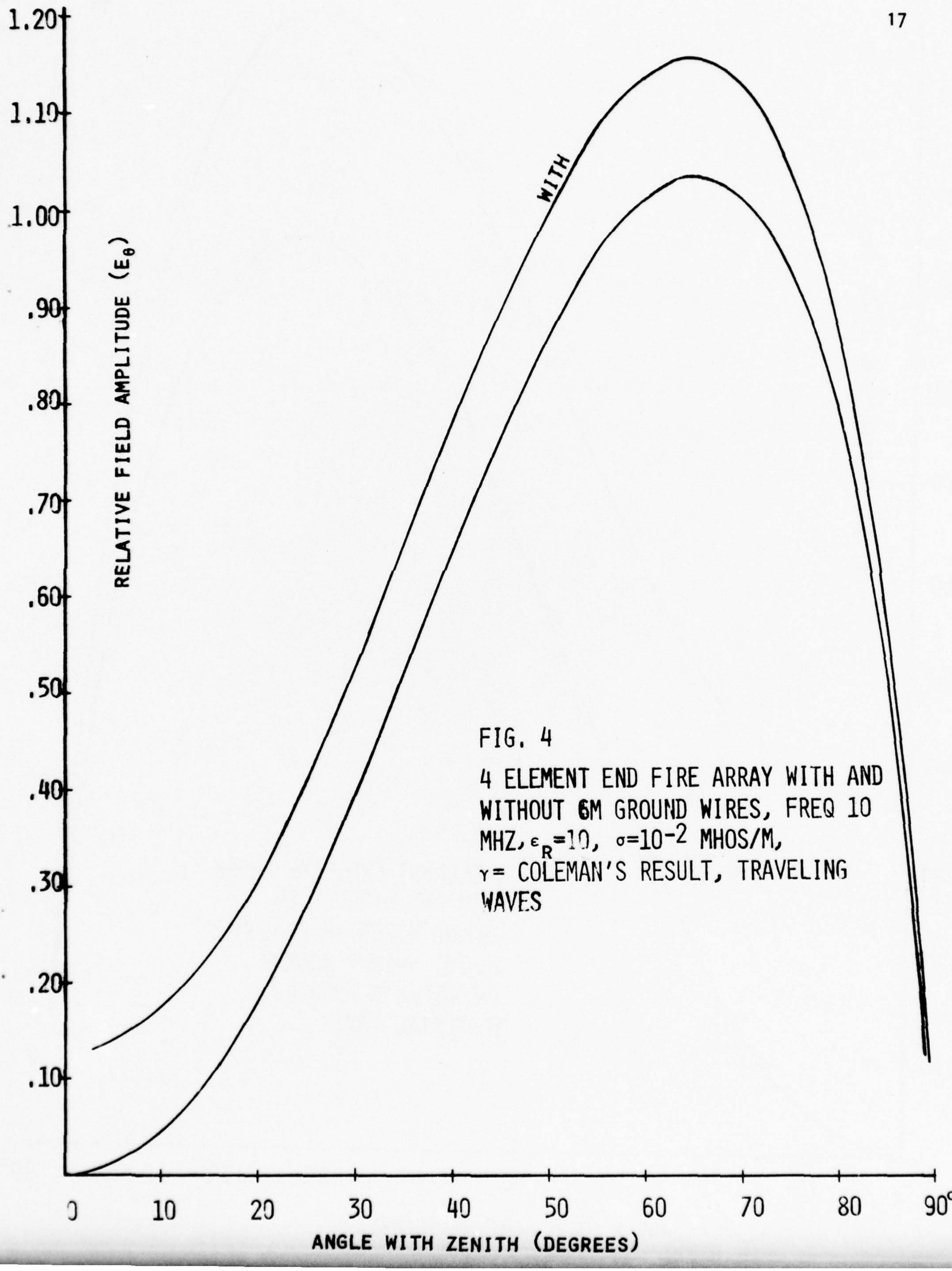


FIG. 4

4 ELEMENT END FIRE ARRAY WITH AND WITHOUT 6M GROUND WIRES, FREQ 10 MHZ, $\epsilon_R=10$, $\sigma=10^{-2}$ MHOS/M, γ= COLEMAN'S RESULT, TRAVELING WAVES

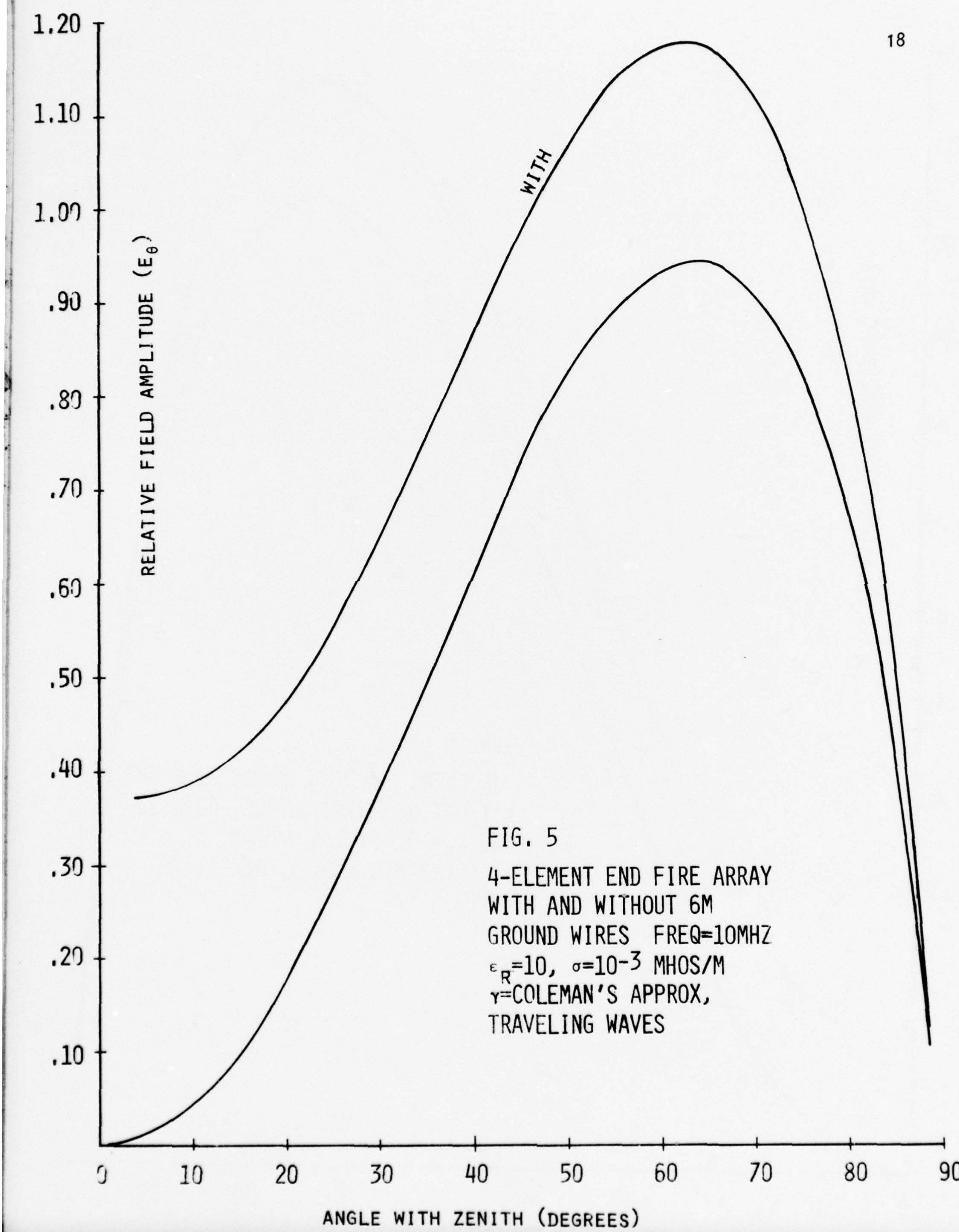


FIG. 5

4-ELEMENT END FIRE ARRAY WITH AND WITHOUT 6M GROUND WIRES FREQ=10MHZ ϵ_R=10, σ=10^{-3} MHOS/M γ=COLEMAN'S APPROX, TRAVELING WAVES

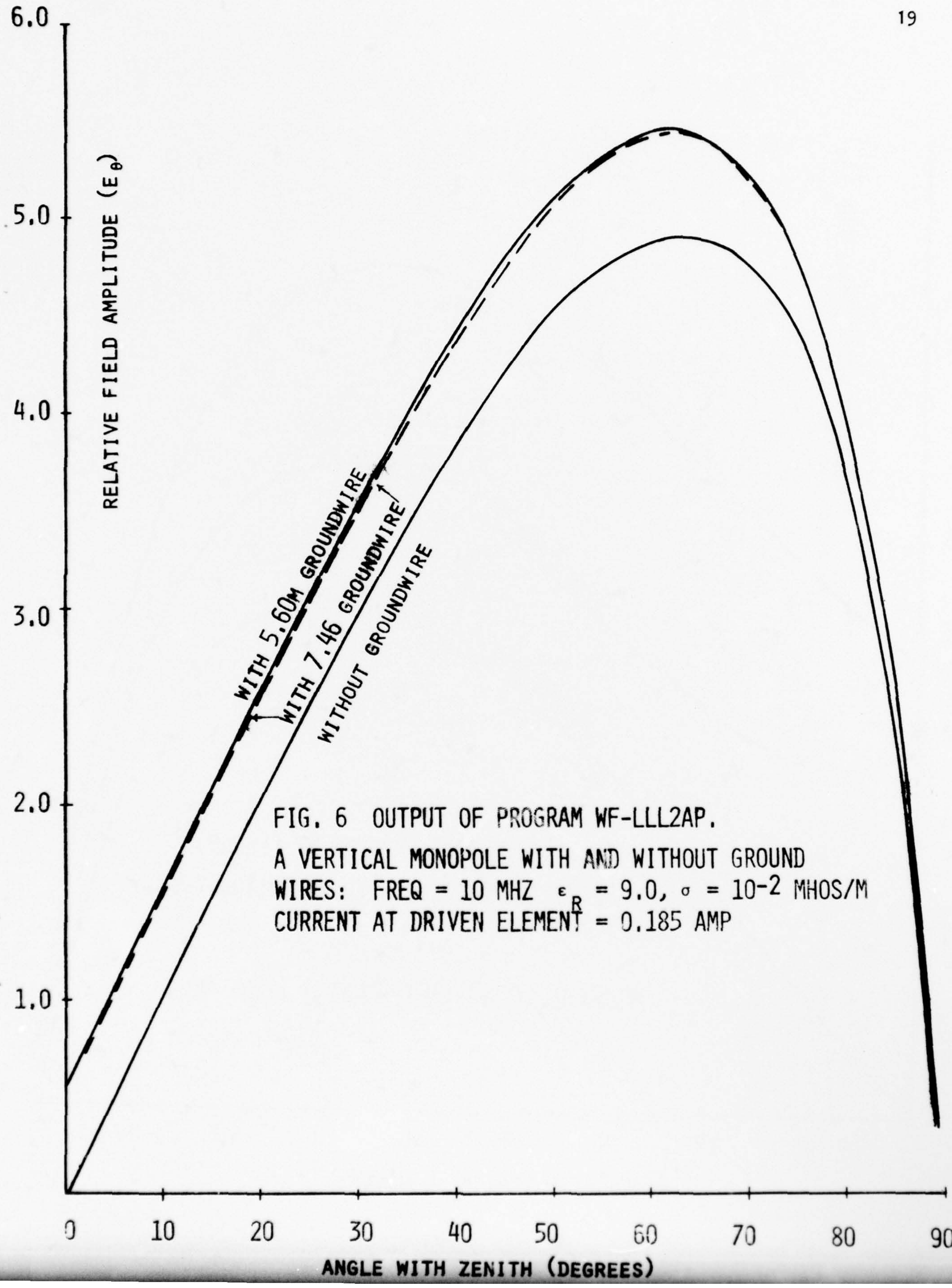


FIG. 6 OUTPUT OF PROGRAM WF-LLL2AP.

A VERTICAL MONOPOLE WITH AND WITHOUT GROUND WIRES: FREQ = 10 MHZ $\epsilon_R = 9.0$, $\sigma = 10^{-2}$ MHOS/M CURRENT AT DRIVEN ELEMENT = 0.185 AMP

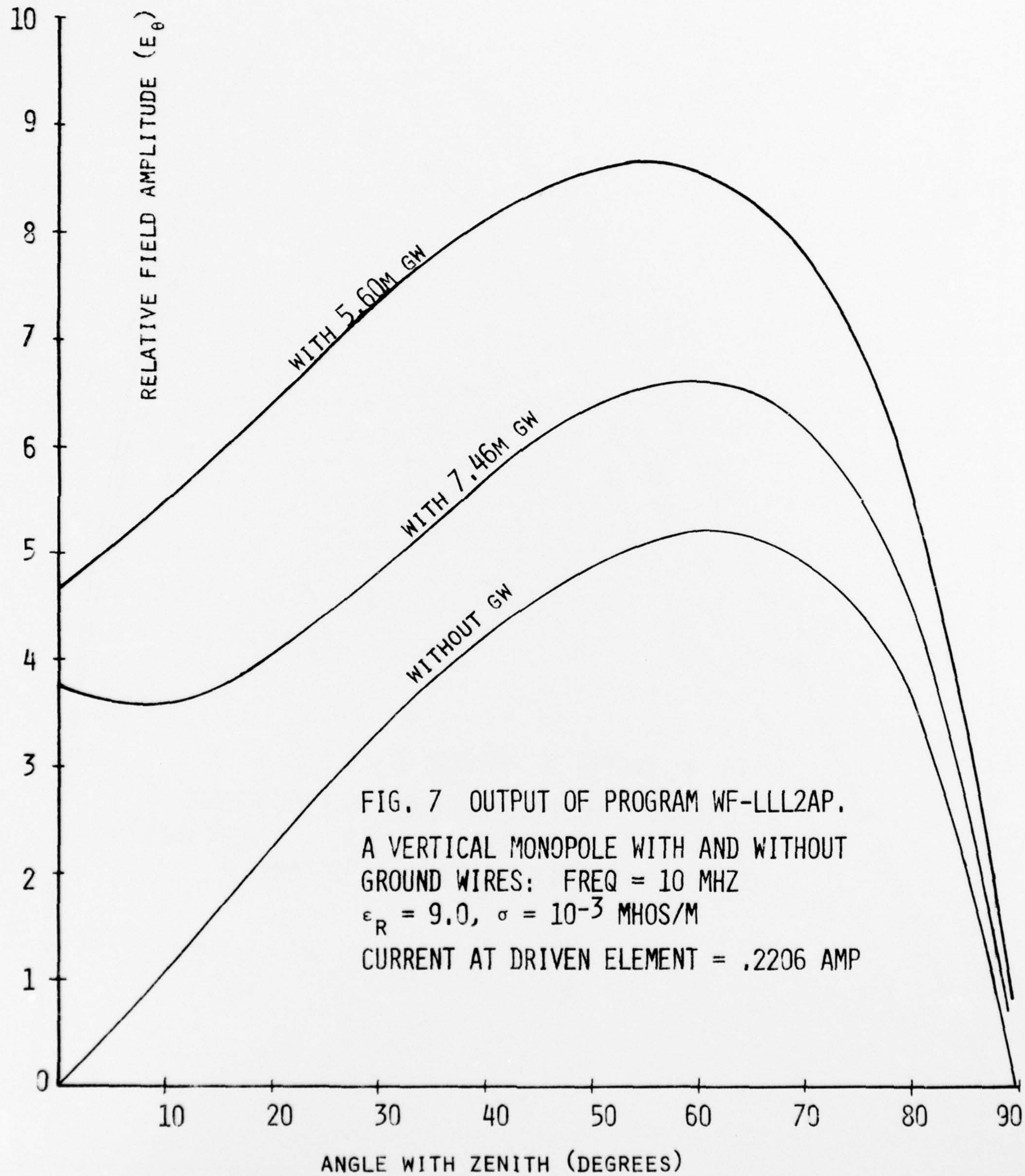


FIG. 7 OUTPUT OF PROGRAM WF-LLL2AP.

A VERTICAL MONOPOLE WITH AND WITHOUT GROUND WIRES: FREQ = 10 MHZ
$\epsilon_R = 9.0$, $\sigma = 10^{-3}$ MHOS/M
CURRENT AT DRIVEN ELEMENT = .2206 AMP

judge from the patterns. For this reason, the balance of data presented does not give the whole radiation pattern but only the enhancement ratio (field magnitude with ground wires/field magnitude with no ground wires) for some particular low angle. For other low angles, the enhancement is not greatly different although it tends to be slightly smaller for the lower angles.

The next Figure, 8, shows enhancement data at 10 MHz, $\theta \approx 75°$, for three different earth conductivities, as a function of the length of the horizontal ground wires (same assumptions as above). These data show that for conductivity of 10 millimhos/m, the enhancement is about as high as it ever will be when the ground wires are 4 to 5 meters long, at which time, the maximum enhancement possible would be at 10%. And in fact, the field enhancement goes down slightly for lengths twice as long as the lengths that give the maximum effect. If the conductivity is 5 millimhos/m, an enhancement of about 15% is possible for ground wires of length about 6 m and with longer wires, the possible enhancement is less, falling to about 10% at a length of about 15 meters. For the still lower conductivity, namely 1 millimho/m, more enhancement is possible, about 20% with a length of 6 meters. However, in the data for 1 millimho/m, a cyclic variation of enhancement level with length is apparent and with a length of about 17 meters, the ground wires actually decrease the low angle radiation by perhaps 7 or 8 percent.

The next Figure, 9, gives the corresponding data for a lower frequency, namely 3 MHz, with the same three values of earth conductivity. These data show that a) at the lower frequency, the possible enhancement is less, being perhaps 3% for 10 millimho/m earth, 6% for 5 millimho/m earth and perhaps 18% for 1 millimho/m earth, b) the ground screen length to accomplish the maximum ehancement is longer, in the vicinity of 19 to 20 meters with the lowest value of earth conductivity.

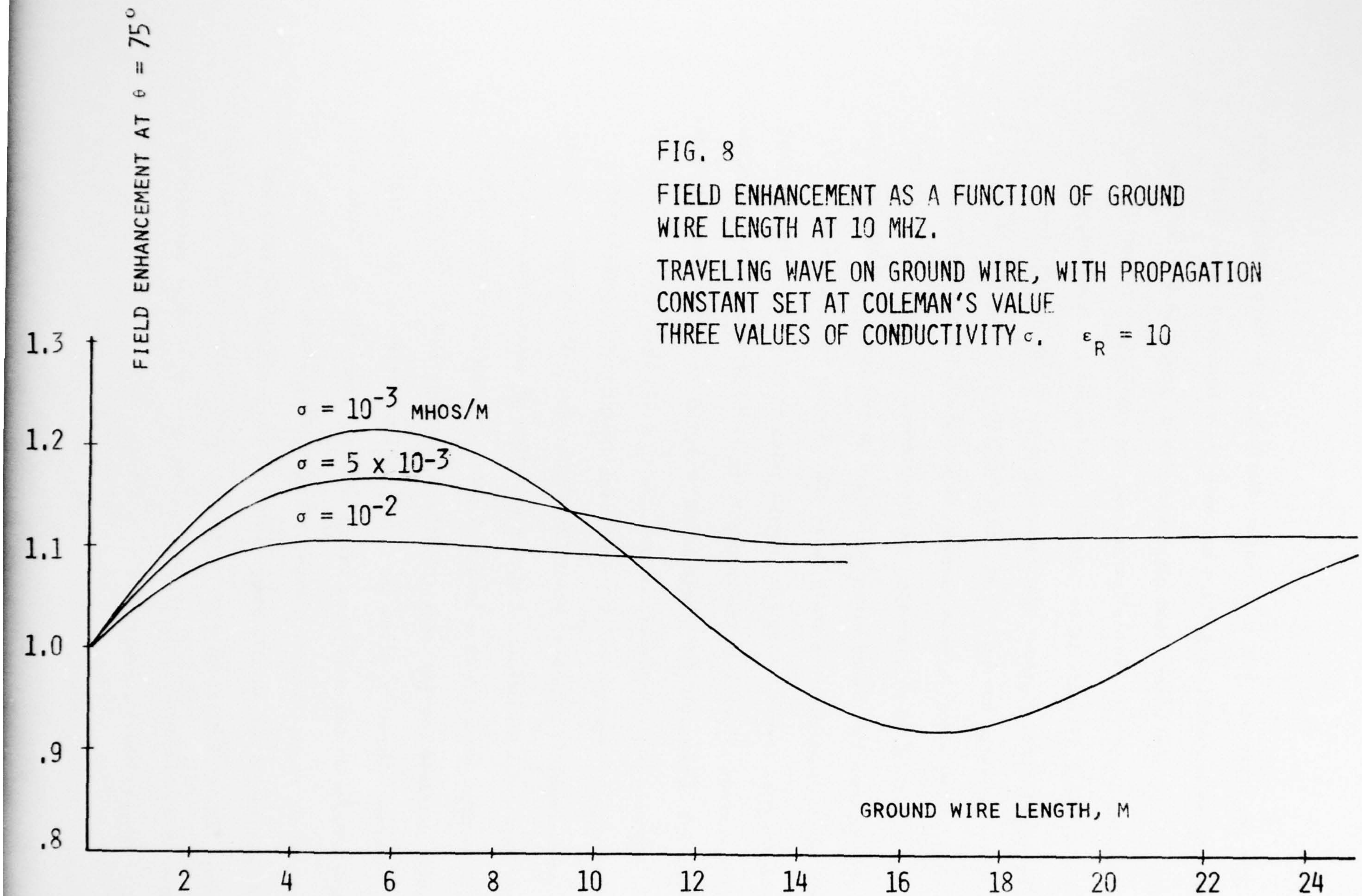


FIG. 8

FIELD ENHANCEMENT AS A FUNCTION OF GROUND WIRE LENGTH AT 10 MHZ.

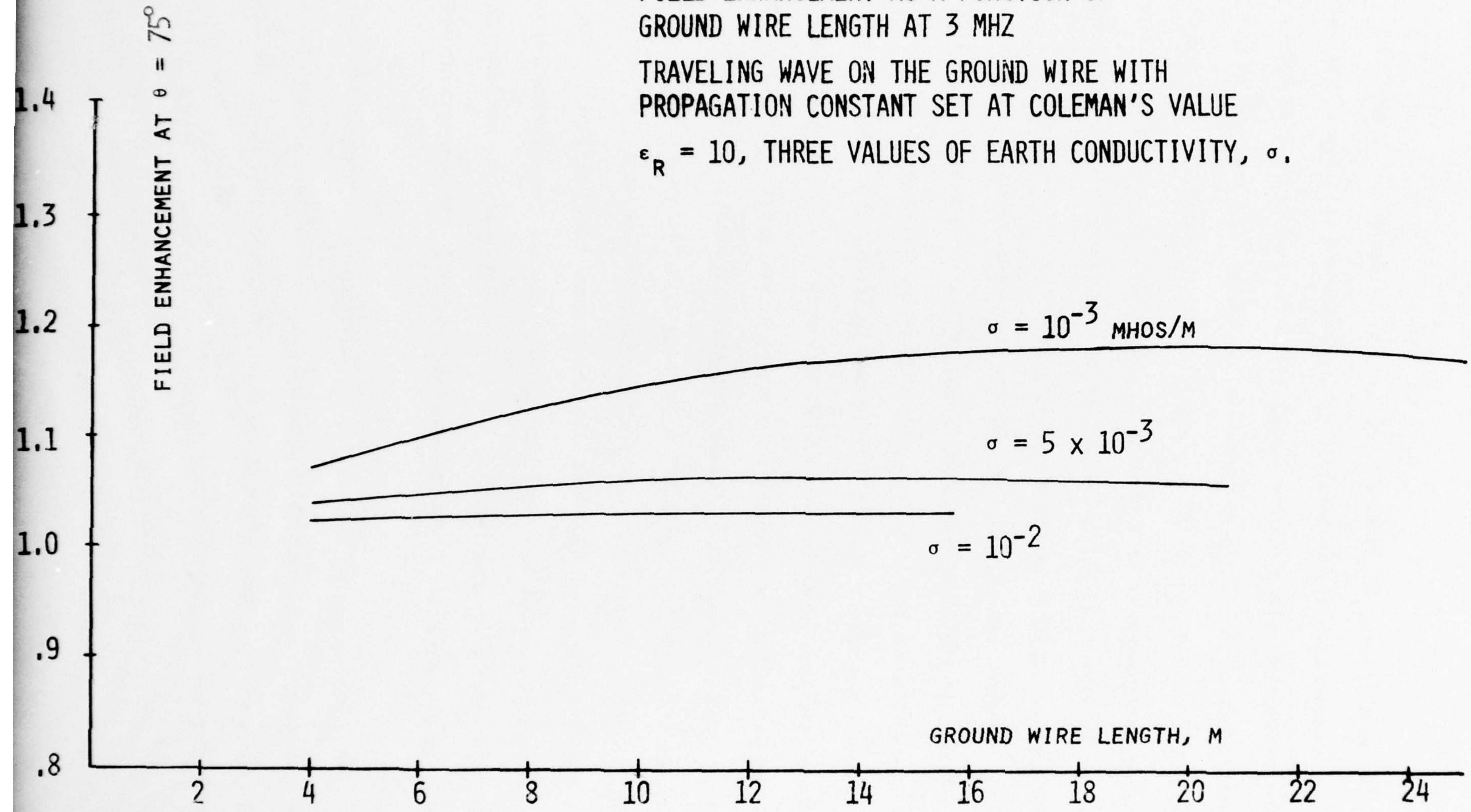
TRAVELING WAVE ON GROUND WIRE, WITH PROPAGATION CONSTANT SET AT COLEMAN'S VALUE
THREE VALUES OF CONDUCTIVITY σ. $\varepsilon_R = 10$

FIG. 9

FIELD ENHANCEMENT AS A FUNCTION OF GROUND WIRE LENGTH AT 3 MHZ

TRAVELING WAVE ON THE GROUND WIRE WITH PROPAGATION CONSTANT SET AT COLEMAN'S VALUE

$\epsilon_R = 10$, THREE VALUES OF EARTH CONDUCTIVITY, σ.



The next Figure, 10, gives the corresponding result for 30 MHz, and of course shows both higher possible enhancements and more drastic variations in the enhancement with length. Again, the variation in the enhancement with length is less with the higher conductivity values. And the data shows that enhancements of about 9 to 19% are possible with conductivity 10^{-2} mhos/meter.

The foregoing data were obtained on the assumption and on imposition of a traveling wave distribution of current on the ground wires, largely having a propagation constant equal to Coleman's approximation p. 10, eqn. 6. Coleman's approximation is valid only for the wires essentially lying in the interface. If the wires have significant insulation around them or if they are somewhat above the interface, the propagation constant changes significantly. To document the effect of such differences in propagation constant, the calculations of the type given in Fig. 10 (30 MHz) were repeated with different (Fig. 11) assumed values for the propagation constant of the traveling wave on the ground wire. A number of different propagation constants were assumed, to show the effect of both a change in the real part (attenuation) and the imaginary part (phase constant) of the propagation constant. Of course, with the higher attenuation constants, the enhancement tends to reach a maximum at some length and to level off thereafter. For the phase constants smaller than that of the Coleman formula, the length that shows the maximum enhancement tends to be longer and if the attenuation is small, the enhancement levels larger.

This dependence of the results on the actual wire propagation constants is important from a practical viewpoint since the positioning and insulation thickness of the ground wires is, at least to some extent, under the control of the system designer.

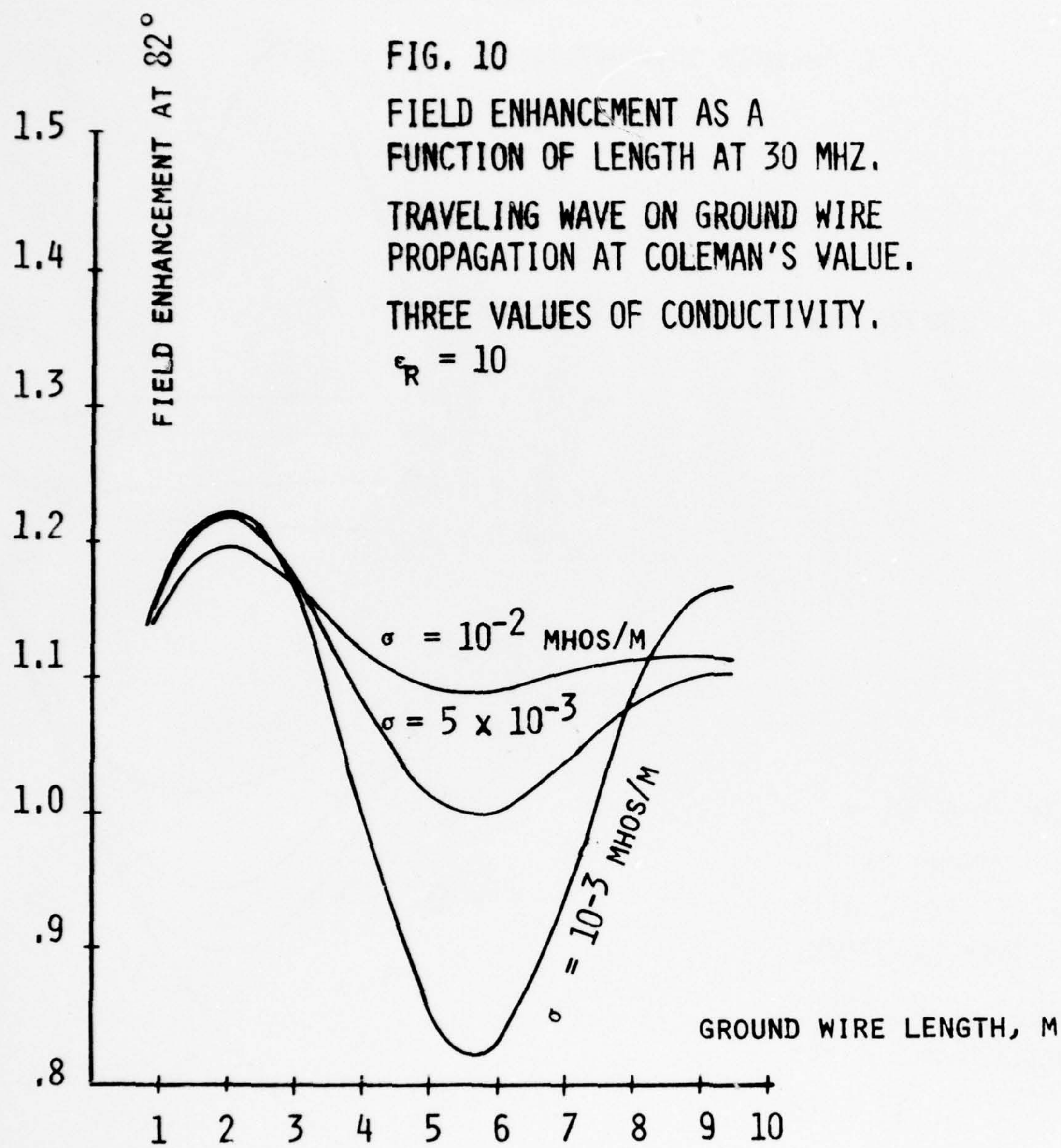


FIG. 10

FIELD ENHANCEMENT AS A FUNCTION OF LENGTH AT 30 MHZ.

TRAVELING WAVE ON GROUND WIRE PROPAGATION AT COLEMAN'S VALUE.

THREE VALUES OF CONDUCTIVITY.

$\epsilon_R = 10$

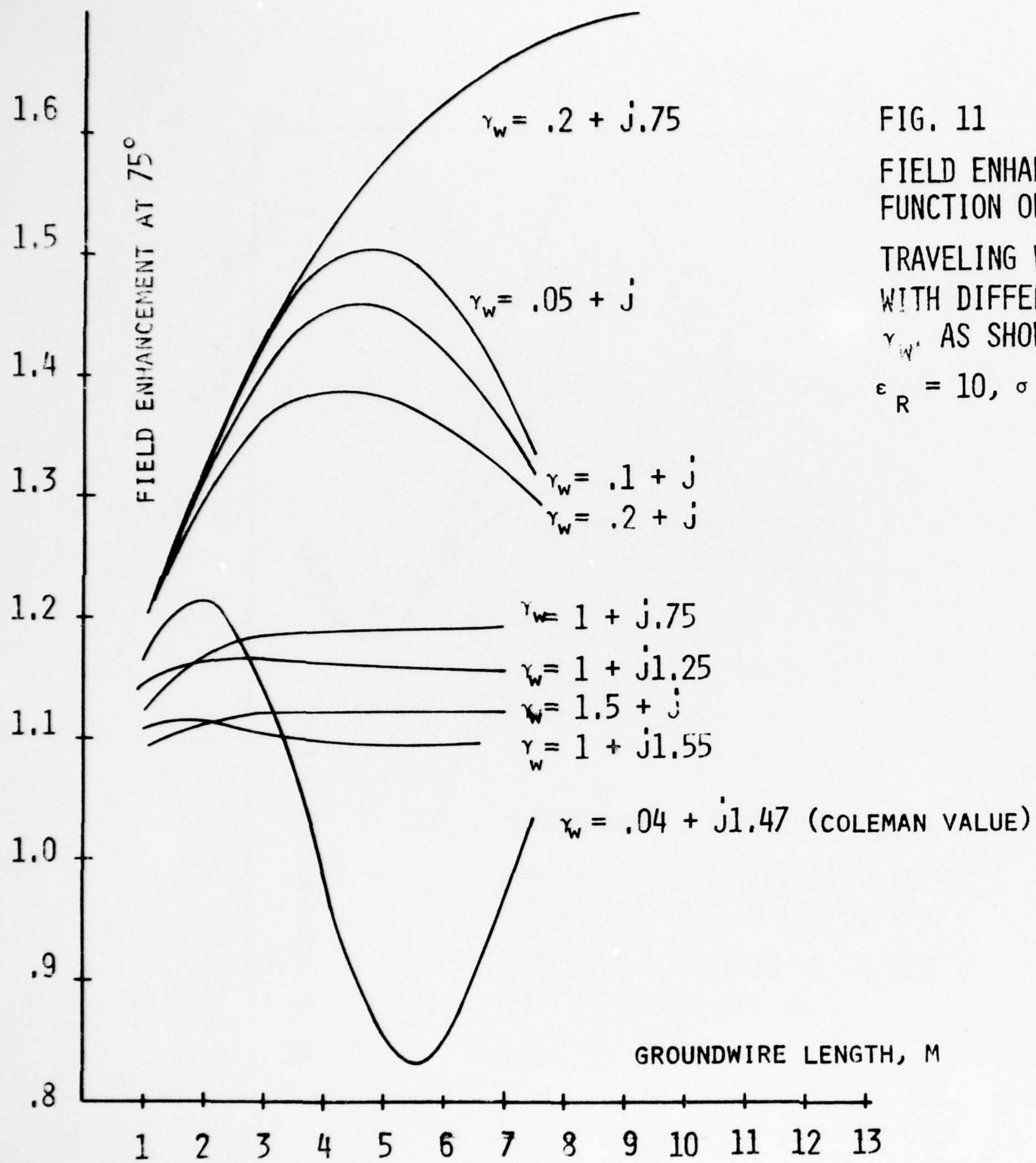


FIG. 11

FIELD ENHANCEMENT AS A FUNCTION OF LENGTH AT 30 MHZ

TRAVELING WAVES ON THE GROUND WIRE WITH DIFFERENT PROPAGATION CONSTANTS, γ_w, AS SHOWN.

$\varepsilon_R = 10$, $\sigma = 10^{-3}$ MHOS/M

Of course, when the WF-LLL2AP program is used in the calculation, no assumption regarding propagation constant or current distribution on the wires is necessary. The program output gives the current values at the center of each segment length used in the calculation. From these data it is possible to obtain an approximation for the propagation constant on the ground wires.

This was done, at a frequency of 10 MHz for two values of conductivity (10^{-3} and 10^{-2} mhos/m) with a wire of 3.77mm radius located 2.1 radii above the interface. In this way it was found that the Coleman result gives a propagation constant that is reasonably close to the apparent propagation constant as inferred from the calculated current distribution.

Another important variable, to some extent under the control of the designer, is the type of termination at the end(s) of the ground wire. The easiest alternative in practice is of course to leave the far end open circuited. But for low conductivity earth and relatively short ground wires, this may not be a desirable alternative. The reason is that the ground wire system may tend to exhibit resonant-impedance conditions for certain lengths. The system effects of such conditions is suggested by the data in Fig. 12; which were obtained by running the computer program, modified WF-LLL2AP. Figure 13 shows similar type effects when the approximate calculation was carried out on the assumption of an open circuit termination on the ground wire. The effect shows up in a striking way because of the internally imposed condition that the total current at the input of the ground wires is equal to the total current at the base of the vertical antennas. The main thing this data shows is that there can be resonant effects on the ground wires under the conditions (low earth conductivity) in which the ground wires are the most important. The input impedance indicated in the WF-LLL2AP program also varies wildly as a function of ground wire length.

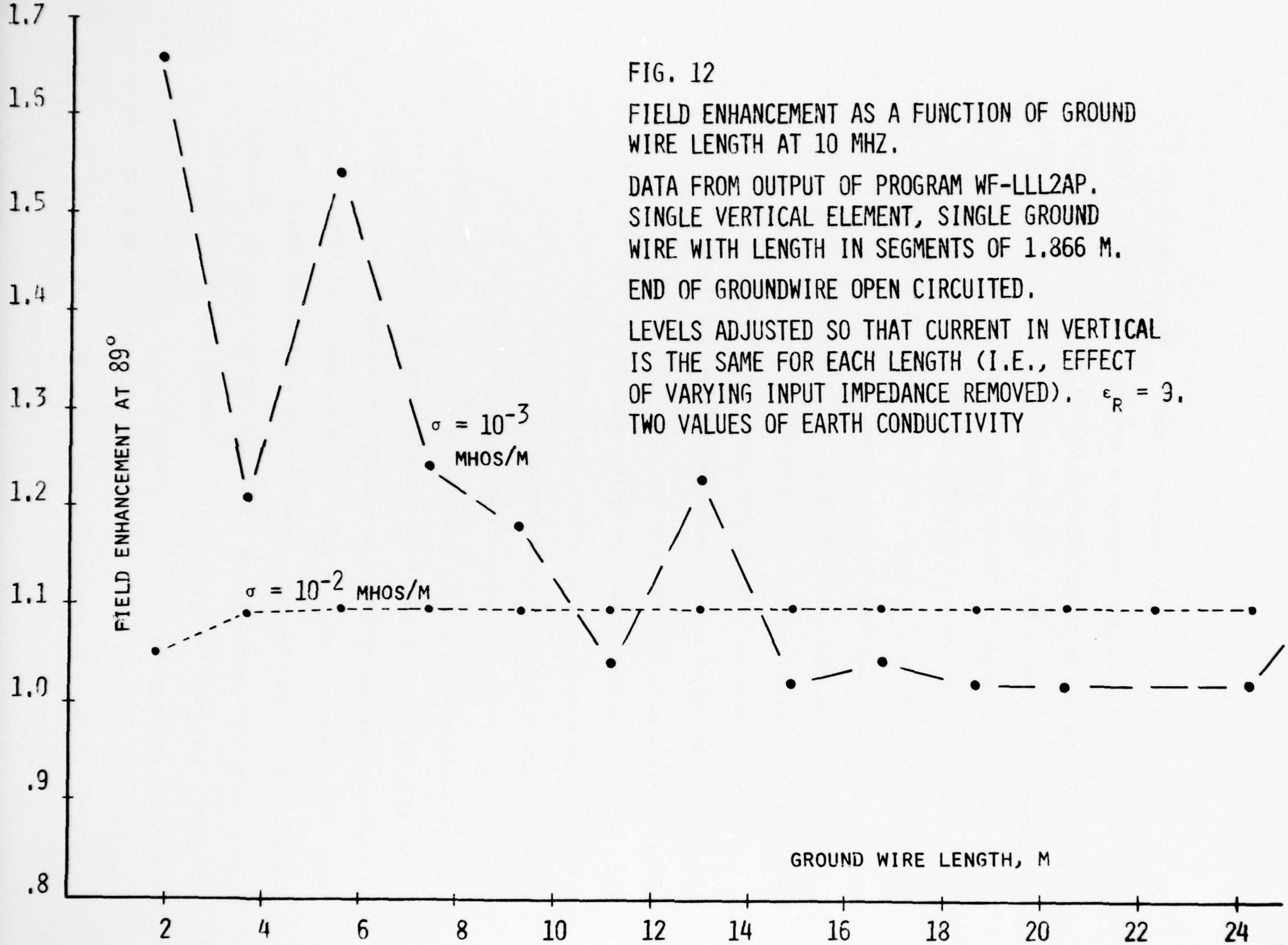


FIG. 12

FIELD ENHANCEMENT AS A FUNCTION OF GROUND WIRE LENGTH AT 10 MHZ.

DATA FROM OUTPUT OF PROGRAM WF-LLL2AP. SINGLE VERTICAL ELEMENT, SINGLE GROUND WIRE WITH LENGTH IN SEGMENTS OF 1.866 M.

END OF GROUNDWIRE OPEN CIRCUITED.

LEVELS ADJUSTED SO THAT CURRENT IN VERTICAL IS THE SAME FOR EACH LENGTH (I.E., EFFECT OF VARYING INPUT IMPEDANCE REMOVED). $\epsilon_R = 3$. TWO VALUES OF EARTH CONDUCTIVITY

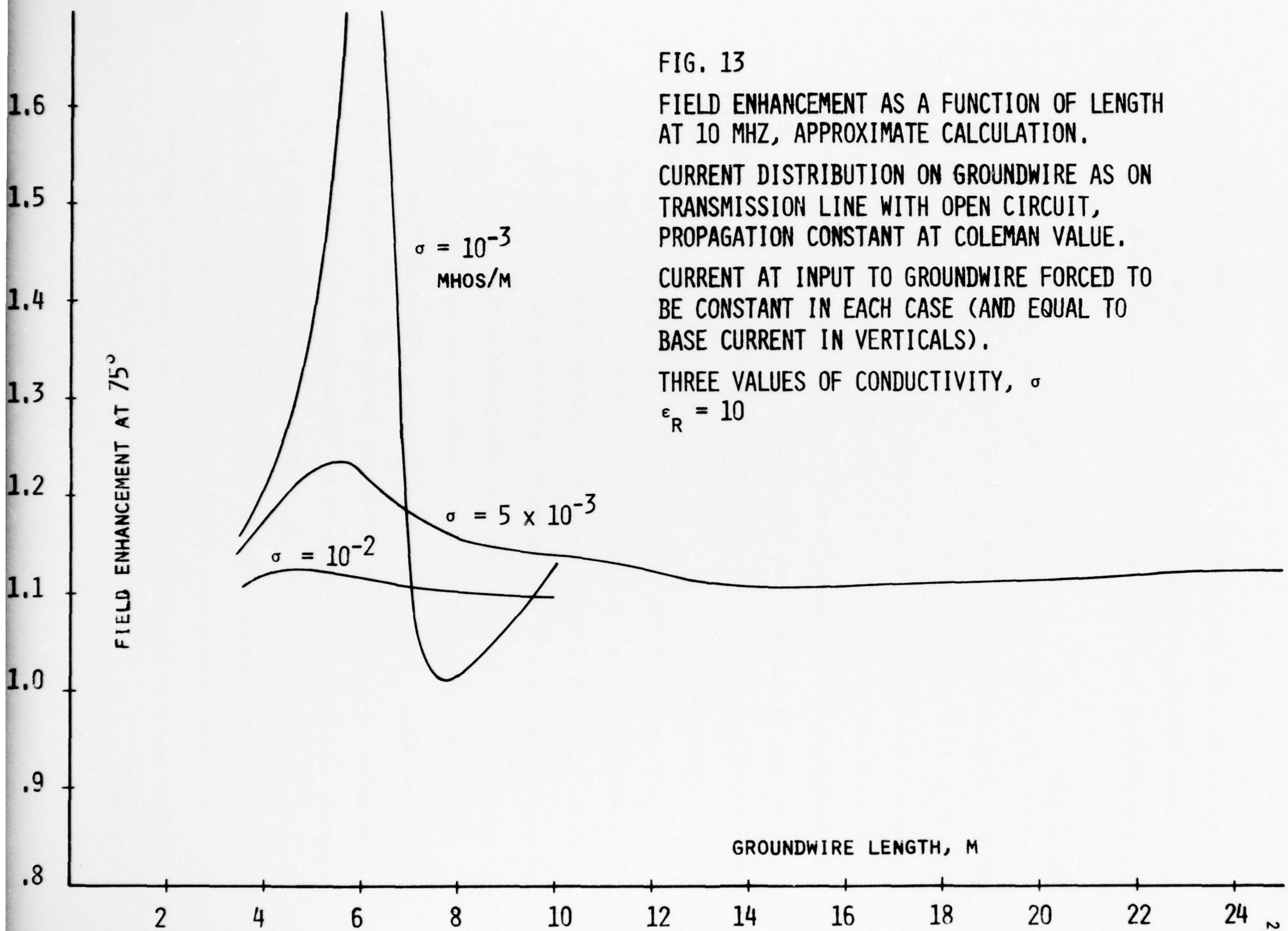


FIG. 13

FIELD ENHANCEMENT AS A FUNCTION OF LENGTH AT 10 MHZ, APPROXIMATE CALCULATION.

CURRENT DISTRIBUTION ON GROUNDWIRE AS ON TRANSMISSION LINE WITH OPEN CIRCUIT, PROPAGATION CONSTANT AT COLEMAN VALUE.

CURRENT AT INPUT TO GROUNDWIRE FORCED TO BE CONSTANT IN EACH CASE (AND EQUAL TO BASE CURRENT IN VERTICALS).

THREE VALUES OF CONDUCTIVITY, σ

$\epsilon_R = 10$

Short circuiting (low impedance) terminations on the ground wires can give similar effects, although in practice, such low impedance terminations would be difficult to achieve in practice if the earth conductivity is small.

Conclusions

After some months of effort, we have concluded that it is possible to employ a method of moments program incorporating a Sommerfeld type of dipole field solution to evaluate the effects of wire ground systems in the HF band. However, with a computer of the type available at Purdue University (CDC 6500) it is so far only marginally so. It is marginal 1) because the evaluation of the "self-impedance" and nearby "mutual impedance" type terms requires so much time and, 2) because of the limitations on the system size imposed by the difficulties and the time required to invert the associated large order system impedance matrix.

By a comparison of the results of the digital computer output with the approximate calculation, we conclude that the assumption of a transmission line like current distribution having a propagation constant calculated in standard fashion (such as Coleman result) can be employed to give useful information on the radiation effects of wire ground systems.

In general, the ground wires may enhance the radiation at all forward and upward angles. However, the amount (and even whether the field is increased or decreased) of the enhancement depends on the ground wire length and propagation constant. For low conductivity earth, resonance effects occur with certain lengths.

For low conductivity earth, enhancements of about 20% appear to be possible, but the results are length and frequency sensitive.

Wire lengths that are long enough or properly terminated so as to give a traveling wave distribution on the ground wires have potential bandwidth advantages.

From a systems point of view, the fact that the designer has some control over the ground wire propagation constant (through insulation thickness) and over the detailed type of grounding connections may be of vital importance to the enhancement of low angle radiation.

On the negative or discouraging side, it must be noted that in no case studied so far can the field enhancement be said to really alter the basic "null at the horizon" effect; for even an increase by a factor of two, which might be possible, still leaves the field near the horizon at quite a small value. Moreover, the patterns show an increase of field strength in the whole forward quadrant and it is not yet clear that the power represented by these field increases comes from power that might otherwise have been dissipated in the earth or whether the additional power would have to be supplied by the transmitter.

References

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2. D. L. Lager and R. J. Lytle, "Fortran subrountines for the numerical evaluation of Sommerfeld integrals unter anderem", Report UCRL-51821, Lawrence Livermore Laboratory, University of California/Livermore, CA (May 1975).

3. W. L. Weeks, Antenna Engineering, McGraw-Hill Book Company (1968).

4. Coleman, B. L. "Propagation of Electromagnetic Disturbance along a Thin Wire in a Horizontally Stratified Medium", Philosophical Magazine, Vol. 41, Series 7, pp. 276-288, 1950.

METRIC SYSTEM

BASE UNITS:

Quantity	Unit	SI Symbol	Formula
length	metre	m	...
mass	kilogram	kg	...
time	second	s	...
electric current	ampere	A	...
thermodynamic temperature	kelvin	K	...
amount of substance	mole	mol	...
luminous intensity	candela	cd	...

SUPPLEMENTARY UNITS:

Quantity	Unit	SI Symbol	Formula
plane angle	radian	rad	...
solid angle	steradian	sr	...

DERIVED UNITS:

Quantity	Unit	SI Symbol	Formula
Acceleration	metre per second squared	...	m/s
activity (of a radioactive source)	disintegration per second	...	(disintegration)/s
angular acceleration	radian per second squared	...	rad/s
angular velocity	radian per second	...	rad/s
area	square metre	...	m
density	kilogram per cubic metre	...	kg/m
electric capacitance	farad	F	A·s/V
electrical conductance	siemens	S	A/V
electric field strength	volt per metre	...	V/m
electric inductance	henry	H	V·s/A
electric potential difference	volt	V	W/A
electric resistance	ohm		V/A
electromotive force	volt	V	W/A
energy	joule	J	N·m
entropy	joule per kelvin	...	J/K
force	newton	N	kg·m/s
frequency	hertz	Hz	(cycle)/s
illuminance	lux	lx	lm/m
luminance	candela per square metre	...	cd/m
luminous flux	lumen	lm	cd·sr
magnetic field strength	ampere per metre	...	A/m
magnetic flux	weber	Wb	V·s
magnetic flux density	tesla	T	Wb/m
magnetomotive force	ampere	A	...
power	watt	W	J/s
pressure	pascal	Pa	N/m
quantity of electricity	coulomb	C	A·s
quantity of heat	joule	J	N·m
radiant intensity	watt per steradian	...	W/sr
specific heat	joule per kilogram-kelvin	...	J/kg·K
stress	pascal	Pa	N/m
thermal conductivity	watt per metre-kelvin	...	W/m·K
velocity	metre per second	...	m/s
viscosity, dynamic	pascal-second	...	Pa·s
viscosity, kinematic	square metre per second	...	m/s
voltage	volt	V	W/A
volume	cubic metre	...	m
wavenumber	reciprocal metre	...	(wave)/m
work	joule	J	N·m

SI PREFIXES:

Multiplication Factors	Prefix	SI Symbol
$1\ 000\ 000\ 000\ 000 = 10^{12}$	tera	T
$1\ 000\ 000\ 000 = 10^{9}$	giga	G
$1\ 000\ 000 = 10^{6}$	mega	M
$1\ 000 = 10^{3}$	kilo	k
$100 = 10^{2}$	hecto*	h
$10 = 10^{1}$	deka*	da
$0.1 = 10^{-1}$	deci*	d
$0.01 = 10^{-2}$	centi*	c
$0.001 = 10^{-3}$	milli	m
$0.000\ 001 = 10^{-6}$	micro	μ
$0.000\ 000\ 001 = 10^{-9}$	nano	n
$0.000\ 000\ 000\ 001 = 10^{-12}$	pico	p
$0.000\ 000\ 000\ 000\ 001 = 10^{-15}$	femto	f
$0.000\ 000\ 000\ 000\ 000\ 001 = 10^{-18}$	atto	a

* To be avoided where possible.